*Detlef Burchard, Dipl.-Ing., Box 14426, Nairobi, Kenya*

# Absolute Calibration of a Noise Source

Semi-conductor noise sources have become common everywhere for determining the noise levels of receivers. As against the tube diodes used earlier, they have the advantage of a very long life and the disadvantage that their excess noise ratio can not be derived from a natural law. They must be individually calibrated!

To this end, a primary standard is required, which can be, for example, a tube diode noise generator.

Should no standard be available, an absolute calibration can still be carried out, which need not be as accurate as that of industrial products.

The present article gives three methods, which differ as to cost and precision. A receiver, a signal generator and a digital voltmeter are virtually all you need.

## 1. DEFINITIONS AND PRINCIPLES

According to the Nyquist formula, the noise output available from an ohmic resistance during tuning is:

$$P_r = k * T * B \qquad (1)$$

k = Boltzmann's constant: $1.38.10^{-23}$Ws/K
B = Band width in Hz
T = Absolute temperature in K

At the reference temperature, 290 K (+ 17° C), therefore, it is $4.10^{-21}$Ws (or W/Hz) and corresponds to -174 dBm/Hz. The "strength" of noise sources, the excess noise ratio, is a dimensionless factor, which indicates how many times higher the available noise output is than the thermally generated noise output, in accordance with equation (1). The German expression which describes this best, although
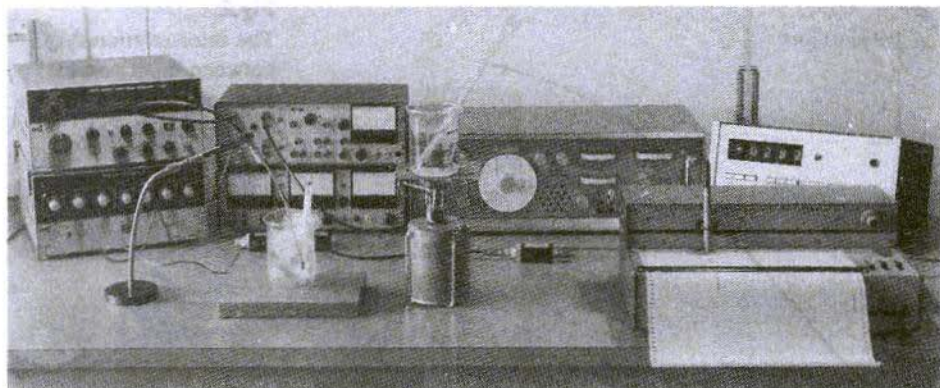
Fig.1: Equipment required for Calibrating a Noise Source:
Rear left to right: signal generator; next bottom, power supply for
semiconductor noise source; on top, multi-function unit containing two
30dB broad-band amplifiers which ensure low-noise input and/or
decoupled intermediate frequency output before or after the laboratory
receiver; a laboratory receiver; finally a digital voltmeter with 10uV
resolution.
Front left to right: home made 50 ohm sensor, home made noise generator
(partially covered by glass), glass containing melting ice, glass containing
boiling water. The chart recorder at the front right is for documentation
purposes only.
The home made measuring head can be seen in front of the receiver.

an unfamiliar one, is the "Rauscherhoehungs-faktor" (noise increase factor). It is calculated in dB, so here I shall describe the expression 10.lg ENR as the noise increase dimension and use ENR/dB as the formula. This happens in an analogous way to additional values, which present themselves both as factors and as logarithmic dimensions (Table 1).

Here we are using only outputs for calculation. So you can mentally put the word "output" before all these expressions. With the logarithmic values, that soon becomes apparent anyway from the 10 in front of the logarithm.

A noise source with noise increase factor 35.482 (ENR/dB = 15.5) can be imagined as an ohmic resistance at a temperature 10,000 K higher -sensor (10,000 + 290) - 290. Because of this interaction, industrially manufactured noise sources are predominantly supplied with a noise increase dimension of 15.5dB.

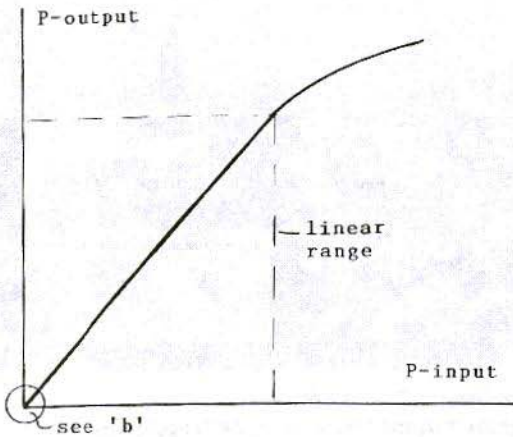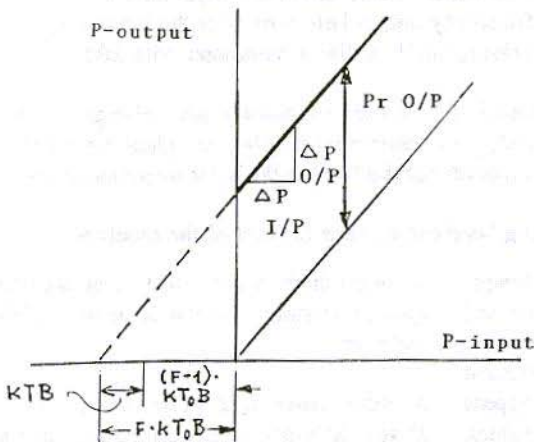| | |
|---|---|
| Noise factor F | Noise dimension F/dB = 10.lg F |
| Change factor Y | Change dimension Y/dB = 10.lg Y |
| Amplification factor G | Amplification dimension G/dB = 10.lg G |
| Damping factor a | Damping dimension a/dB = 10.lg a |
| Reverse damping $a_r$ | Reverse damping dimension $a_r$/dB = 10.lg $a_r$ |

Table 1

Fig.2:
The inter-relationship
between input power
and output power for a
linear amplifier at:
a: high level
b: low level

P-output

linear
range

P-input

a

see 'b'

P-output

Pr O/P

$\Delta P$
O/P

$\Delta P$

I/P

P-input

b

kTB

$(F{-}1) \cdot$
$kT_0B$

$F \cdot kT_0B$

In spite of the enormous apparent increase in temperature, the noise output thus generated is still so small that it must be amplified before it can be displayed using the normal output meters at, for example, - 40dBm. If the band width of the amplifier is 1Hz, then we need approximately 120dB amplification, if 1 kHz/ 1 MHz/1 GHz, then 90/60/30 are the figures required. Not until 1 THz would there be sufficient output to show up on the output meter referred to earlier. Unfortunately there are no broad-band noise sources or output meters of this nature. Moreover, for reasons of

quantum mechanics, formula (1) applies accurately enough only up to approximately 100 GHz. On the other hand, we are not normally interested in the average noise increase dimension over a large frequency range, but rather within quite specific narrow frequency bands. A flat-line pattern can be obtained over a large frequency range. This is advantageous, but not absolutely essential.

Amplifiers have internal noise sources of the same order of magnitude as the noise sources to be calibrated. So if amplifiers always have

to be used to calibrate a noise source, then we must initially busy ourselves with the amplifier noise levels. And here we are including as amplifiers those with single frequency conversion or multi-frequency conversion, which have a defined band width in the frequency plane and which also have defined amplification, i.e: a linear relationship between the output power and the input power.

This relationship is shown in Fig.2a. The curve does not show one important detail at the origin. If we draw the zero point area on a larger scale (Fig.2b), it then becomes clear that the curve does not meet the zero point, but meets the ordinate at a positive value. The noise output power, PROutput is the noise output given out by the source resistance at the input and amplified in the amplifier, plus the noise output generated in the amplifier itself. Without any further consideration of where the noise source within the amplifier is concealed, we can assign its effect to the input, i.e. we can act as if this noise output were also subjected to the complete amplification. The noise factor of the amplifier is a dimensionless number, which gives the factor by which the noise of the source resistance is increased in accordance with equation (1). But since it can be measured only at the output, the determination equation

$$F = \frac{P_{R\,Ausgang}}{P_{R\,Quelle} \cdot B \cdot G} \qquad (2)$$

applies, or else

F/dB =
$P_{R\,Ausg}$/dBm + 174 dBm − 10 · lg B/Hz − G/dB
(2a)

Only in special cases (low-frequency amplifiers) can the source resistance be made zero, so that the amplifier noise can be measured on its own. In the high-frequency range, all active and passive components alter their properties if the system resistance (usually 50 ohms) is deviated from.

Should the noise output of a noise source now be fed to the amplifier input, then the output power increases by the change factor, Y. The relationship existing between the ENR, F and Y is as follows:

$$ENR = F \cdot (Y - 1) \quad \text{or} \qquad (3)$$
$$ENR/dB = F/dB + 10 \cdot lg\,(Y - 1) \qquad (3a)$$

This formula is normally used to determine F, with the help of a noise source. Naturally, it can also be used to determine the ENR if an amplifier with a known noise factor is used. If the ENR is known, then conclusions can be drawn about the amplification from the measured increase in the output power and vice versa. It is very frequently true, as indicated in Fig.2b, that the amplification is the quotient obtained from the output power increase and the input power increase. In a completely linear area, we can measure an input power increase of any size as a corresponding output power increase. Thus we do not need particularly low-noise amplifiers to calibrate a noise source. We need only a high-resolution power meter.

A power meter of this type operates through a diode measuring head and a digital display. Anyone who has existing apparatus from HP, Marconi or R & S is already over-equipped. Because in fact we shall predominantly be measuring power ratios, absolute calibration is not required. Vieland-type equipment (5, 7) is also very suitable. A diode measuring head in front of a high- resolution digital voltmeter (10uV or better) is suitable, as is one of the many DVM models with 3.3 places and a minimum measuring range of 200.0mV, to which a chopper amplifier is connected in series, in accordance with Fig.3. The diagram

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also shows how the zero connections must be effected so that noise loops in coaxial cables and thermal stresses do not have any effect on the rectifier.

The diode measuring head operates in accordance with a natural law which we can rely on implicitly at rectified voltages of up to 10mV. As Burchard (1) has demonstrated, the rectified voltage is as follows:

$$U_{Richt} = \frac{1}{2} \frac{\hat{U}^2}{U_T}$$

$$\hat{U} < 2 \cdot U_T$$

$$U_T \approx 40 \text{ mV (AA 119)} \qquad (4)$$

A peak factor of 8 is sufficient for the correct evaluation of noise with Rayleigh distribution (excess probability for 1/2,000 of time). This measuring head then has an input power range of between - 54 and - 21 dBm (rectified voltage of 5uV to 10mV). A doubled reading indicates a doubling of the input power. An increase by a small amount - and here it is irrelevant whether the previous value was low or high - indicates that the input power has increased by an equally small amount. The determination of a change factor, Y, becomes particularly easy. It is simply the quotient of the two figures indicated.

# 2.
# NOISE GENERATORS

Home-made models can be designed like those of Ulbricht (4), Fleckner (2) or Rohde (3). What is common to all of them is that current is sent through a semi-conductor junction with zener or avalanche properties. If the current can be switched on and off, then we have a cycled noise source, the temperature of which can be converted from the reference temperature to a much higher one. Unfortunately, the differential diode resistance changes considerably at the same time. A good match with the system resistance can be obtained only by using damping elements. The desired noise increase factor can be set by the selection of the damping factor. All the noise sources referred to above initially give ENR at 35dB, and can therefore be reduced to the desired value by means of 20dB damping. This simultaneously increases the structural return loss in the desired fashion to approximately 40dB.

There is an optimal current for each example of a noise- generating PN junction. The semi-conductor noise source is not adjustable as is the case for a tube diode. It more closely
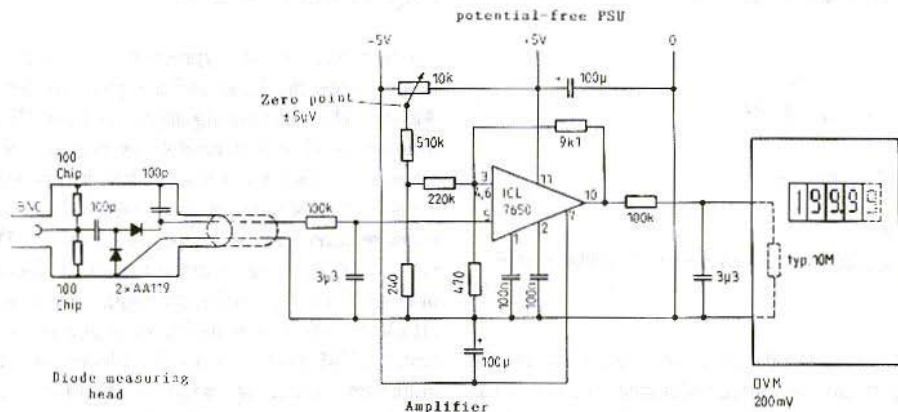


Fig.3:    Circuit for a high-resolution home made Power Indicator

resembles a noise-generating gas discharge, and there is a model which accounts for the noise using an inter-crystalline plasma in the semi-conductor. For this reason, the optimal current must first be found for any home-made noise source.

A variable-frequency receiver can be used to establish whether the noise is white, and therefore independent of frequency. Look for the current level at which this requirement is best met. In principle, this can be done using the laboratory receiver in Fig.1, although it is a laborious task.

Because documentation is easier, a spectrum analyser was used behind the 30 dB amplifier, and thus we obtained Fig.4. With this semi-conductor example (BF324A), a current of 200uA was revealed as correct (Fig.4b). The value is not critical, +/- 20% can change the picture.

However, if the current is tripled, or reduced to a quarter (Fig's.4a and 4c), there is a larger spectral power density, although it can scarcely be called white. The reduction above 600 MHz is because of the wide-band amplifier, which has an NE5205 in the input

This analyser reading should not be over-valued, because the errors can be consider-able. The frequency level of the preamplifier (without which most analysers will in general not give a picture which can be evaluated) is a contributing factor, as are those of the calibration divider and the first mixer, as well as the conversion error of the logarithmic converter. You can't expect much better than 1 to 2dB!

The analyser or receiver used, however, makes possible an initial estimate of the noise increase dimension! Here we are using method 1.

# 3.
# ABSOLUTE CALIBRATION

## 3.1 The noise factor method

If you have a receiver with a known noise dimension and a determined change factor, Y, then you can also calculate ENR using equation (3). The example below, together with all following examples, refers to the structure shown in Fig.1 at 150 MHz.

From Fig.4b, we read off Y/dB = 10. The amplifier has a typical noise dimension of 6.5 dB. Y - 1 is 9, so that finally we obtain ENR/dB = 16.

The procedure is permissible if the noise dimension of the amplifier is determined by a law of nature. This is largely the case with the NE5205. Its noise predominantly comes about in internal feedback resistances. This also applies to many other gain-blocks.

If the change factor, Y, is determined using a power indicator, then this can be done with great accuracy. The measurement error here depends almost entirely on the accuracy with which the noise dimension is known.

## 3.2 The substitution method

The noise source to be calibrated generates an output increase at the amplifier output which is given by the input power, PR, as per equation (1), multiplied by ENR and the amplification factor. PR contains the effective noise band width of the amplifier, which in general deviates from the -3 dB band width. Using the substitution method, another, known signal source is now connected to the input after the noise source to be calibrated, and the output power increase caused by this is measured.
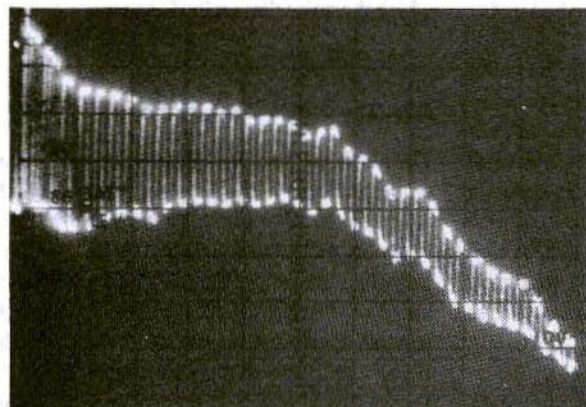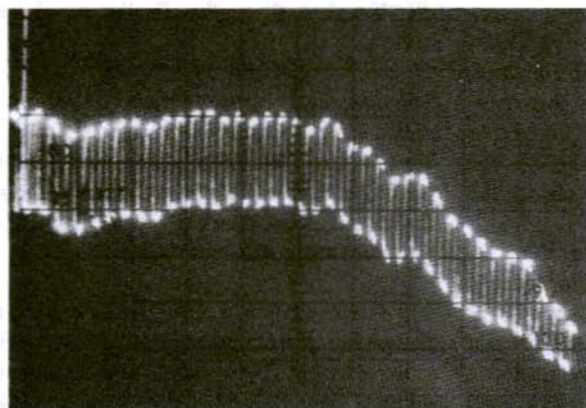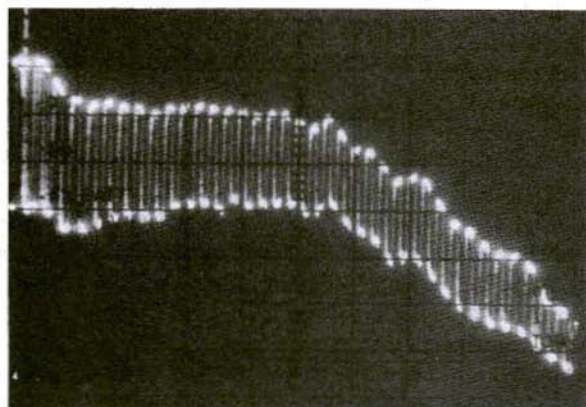
Fig.4:
Spectral power density of a semiconductor noise generator at various operating currents:
a: 650uA
b: 200u;
c: 50uA
X: 0-1 GHz
(100 MHz/div)
Y: -135...-95dBm
(5dB/div)
Bandwidth: 200 kHz
Video Filter: 100 Hz
Sweep time: 3 seconds
Switching frequency of noise source: 15 Hz

If the second source is likewise a noise generator, then we do not need to know the noise band width. From the two change factors and the known noise increase factor of the comparison source, we can immediately calculate:

$$ENR = \frac{Y_1}{Y_2} \cdot ENR_2 \quad oder \qquad (5)$$

$$ENR/dB = 10 \cdot lg \frac{Y_1}{Y_2} + ENR_2/dB \qquad (5a)$$

It is possible to simplify the measurement if the comparison source has adjustable power (tube noise generator). If we actually make Y2 equal to Y1, then we have made ENR = ENR2 and can read the value off directly from the comparison generator.

If the comparison generator is a signal generator with a sine signal, then it is tuned to the maximum value of the response curve of the amplifier. ENR can then be calculated as:

$$ENR = \frac{Y_1}{Y_2} \cdot \frac{P_{Gen}}{k \cdot T_0 \cdot B_{eff}} \quad oder \qquad (6)$$

$$ENR/dB = 10 \cdot lg \frac{Y_1}{Y_2} + P_{Gen/dBm}$$

$$+ \ 174 \ dBm - 10 \ lg \ B_{eff}/Hz \qquad (6a)$$

In this equation, $B_{eff}$ is initially unknown, but can immediately be detected using the equipment available. The effective noise band width is defined by:

$$B_{eff} = \frac{1}{G_{max}} \int_0^\infty G(f) \, df \qquad (7)$$

In practise, continuous integration is replaced by step-by-step formation of totals,

$$B_{eff} = \frac{1}{G_{max}} \cdot \sum_0^\infty G_n \cdot \triangle f, \qquad (7a)$$

where the stages have to be measured sufficiently precisely. The integration limits, 0 and

infinity, indicate that all secondary reception points are included. It can thus be assumed that they are being damped by 30 dB or more, otherwise the noise power received there would be making a noticeable contribution.

Fig.5 demonstrates the experimental reception of the noise band width. The signal generator is changed to the frequency step by step, beginning with frequencies at which the basic noises have not yet been noticeably increased, and ending with the same criterion. The sum of all increases, $G_n$, over the background noise, divided by the highest value, $G_{max}$, and multiplied by a frequency stage, gives the noise band width.

The filter of the laboratory receiver measured here is a Bessel filter with a nominal band width of 200 kHz. The noise band width was measured at 258 kHz. We paid attention to the fact that measurement could be carried out using a very low input power -102dBm (approximately 11dB above the background noise) and was nevertheless most accurate.

Fig.6 is intended to make substitution using the signal generator clearer still. First, a stepped curve was plotted, with the generator power set at values from -107 to -103dBm, and then the noise generator to be calibrated was connected. Now, by interpolation between the steps, the noise source can be determined as equivalent to -104.8dBm or calculated by means of the change factor, Y (on the basis of the -104dBm step, we obtain -104.88dBm, but on the basis of the -105dBm step it is -104.83dBm, which points to 1dB stages which are not quite correct). Using equation (6a), we obtain ENR/dB = 15.

Measurement errors are predominantly due to possible inaccuracies in the output power of the signal generator. For the model used here, the error can amount to +/-2dB. It is made up of the error in the volume control plus that of
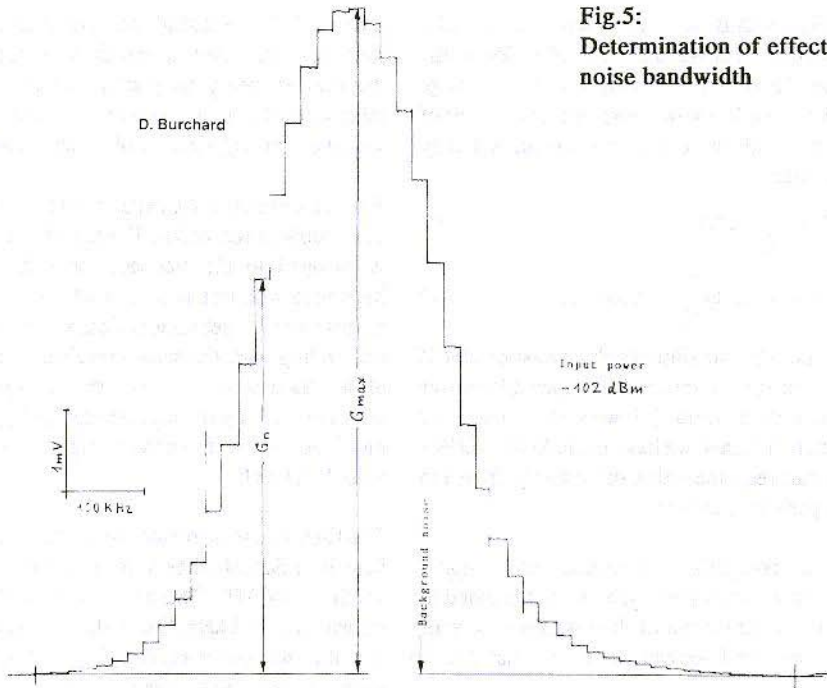
Fig.5:
Determination of effective
noise bandwidth

the calibration divider. If it is desired to obtain better accuracy, then an external calibration divider of greater accuracy can help, with simultaneous measurement of the input power supplied to it (feed-through measuring head).

## 3.3 The temperature method

Until now, we had assumed that the noise resistance was at the reference temperature, 290K. That is certainly not always true, because the apparatus also warms up. However, the error caused by this is normally small in comparison to all the others. We now elevate this effect into a principle! The line entered on the abscissa in Fig.2, to the left of the zero point, $F \cdot k \cdot T_0 \cdot B$, is made up of the thermal noise levels from the source resistance $k \cdot T \cdot B$ and the amplifier noise. We can change the first section if we expose the source resistance to different temperatures.

Such warm-cold standards are commercially available. We are assisted by a metal film resistor, which is exposed to the temperature of melting ice (273K) or boiling water (approximately 373K, depending on the air pressure). Its noise power then changes to approximately $0.34k \cdot T0$.

The power indicator, as per Fig.3, can resolve this small change without problems!

Next I soldered on a sensor, as sketched in Fig.7. It must be water-tight, so that the resistances remain dry, and be able to withstand pressure changes which arise internally at the different temperatures. Multiple dipping in polyurethane lacquer is sufficient. Pressure changes obviously balance out due to the short piece of coaxial cable and the ventilation in the plug. A structural return loss measurement using a rho-bridge (Fig.8) can determine whether it remains tight. Here, a is the
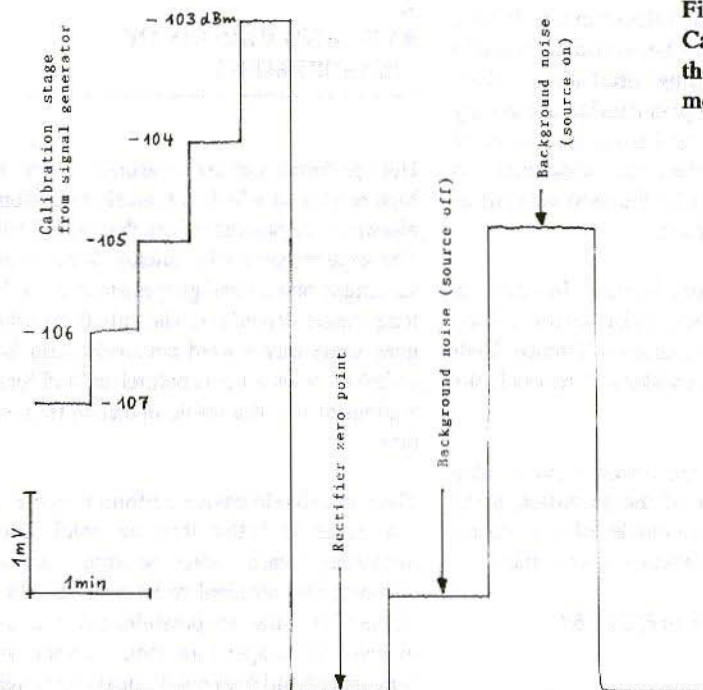
Fig.6:
Calibrations using
the substitution
method

reference level for $a_r = 1(a_r/dB = 0)$, b is the $a_r/dB$ curve for a good sensor and c indicates that a small amount of water has penetrated. Should the resistances be completely under water, then $a_r$ is already measured at under 200 MHz at 10dB, and the sensor has become unusable.

The actual measurement is carried out as per Fig.9. The amplification selected for the rig is as high as possible, and the background noise should almost reach maximum recording level on the power indicator. The difference between the readings from the cold and the warm sensor is then determined. In Fig.9, it reflects a temperature difference of 95K (in Nairobi, because of the height, water boils at 95 degrees C !), i.e: 0.328k . T0. This completes the calibration of the display.

Then the noise generator to be calibrated is connected up. ENR/dB = 15.5 is too high here,

and a defined damping element must be inserted in order to get close to the calibration power change. Here we used such an element, with 19.8dB available, measured accurately. When the noise source is switched on, together with the damping, the reading goes high, to 0.398 k . T0, with a change factor of Y = 0.398. The noise increase dimension before damping, then, is:

$$ENR/dB = 10 \cdot lg\ Y + a/dB, \qquad (8)$$

which here is 15.8dB.

The measurement error can be kept low if the temperature difference is determined as accurately as possible (use a thermometer) and the statistical measurement variations are kept small. Because of the high amplification, the final positions of the display can change at random, and a further increase in the time constants for the integration elements in Fig.3

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may be indicated. The values can then be read off with more security, but measurements take longer. The remaining oscillation (here +/-10uV) can be comprehended as uncertainty of measurement. Should the measured value itself be 160uV, then the uncertainty is +/-0.27dB. This is by far the most accurate of the procedures proposed.

Improvements can still be made by using an amplifier with a lower noise factor and by having a wider temperature difference. Both are intended to obtain wider differences in the reading.

And we can now obtain a more precise value for the noise factor of the amplifier, using Fig.9. From the magnitude of the values recorded or the numbers noted, we obtain:

$$F = 4.6 \text{ or } F/dB = 6.6$$

# 4.
# AVOIDING ERRORS OF MEASUREMENT

Here performances are measured with a very high resolution which is scarcely to be found elsewhere in practise (finer than +/-0.01dB). The experimenter will quickly discover that all sample objects and gauges are more or less temperature dependent. The only thing which goes some way toward countering this is to switch on several hours beforehand and let the equipment run at a stable operating temperature.

The room should have a uniform temperature - a cellar is better than an attic! Diode measuring heads, noise sources, damping elements and terminal resistances should be touched as little as possible, since a 10K increase in temperature from contact with someone's hand does not dissipate itself again completely until half an hour later!
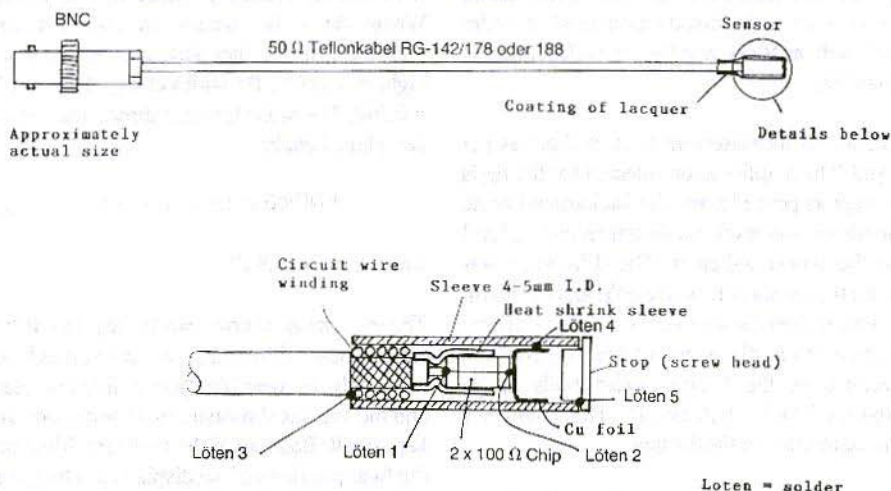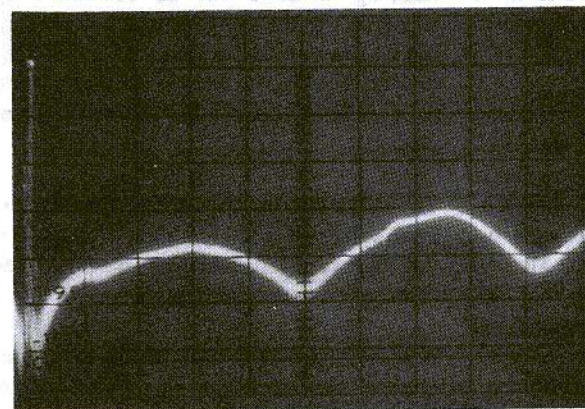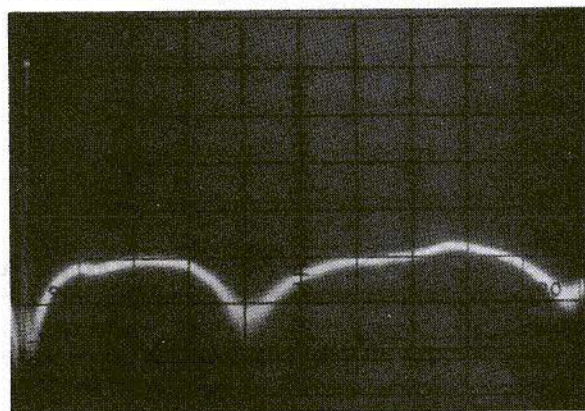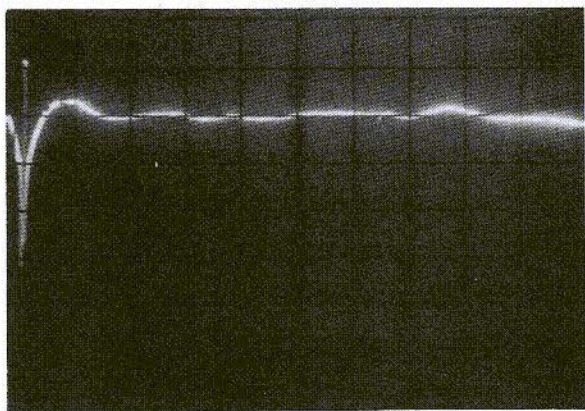


Fig.7:    50 ohm Sensor for temperature method

Fig.8:
Structural return loss of a
50 ohm Sensor measured
using a rho-bridge.
a:  ref. line for $a_r$ = 0dB
b:  measurement curve for
    good sensor
c:  after slight water
    penetration
X:  0...500 MHz
    (50 MHz/div)
Y:  10dB/div
    (ref. line see 'a')
Bandwidth:      200 kHz
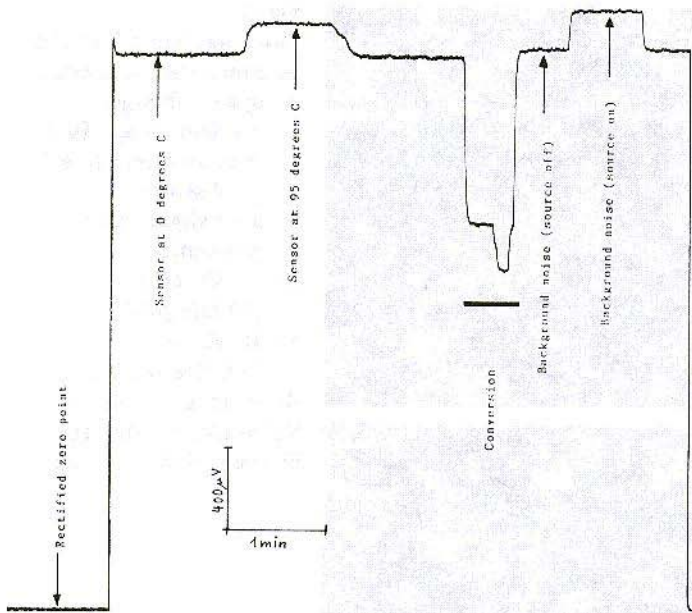Video filter:    10 kHz
Sweep period: 1 second

**Fig.9:**
**Calibration using the temperature method**

All sources and terminals should have high structural return loss values. The absolute bottom limit is 20dB. Here 1% of the power is already being reflected, and an error of measurement of the same size can arise. Figures can be read off from Fleckner (2). If you have a minimum of $a_r = 30$ dB, then there are no problems.

RG cables are good enough, but their damping is not reliable. A section 0.5M long with BNC connections has about 0.15dB of damping at 150 MHz. Make sure that 75 ohm pin-and-socket connectors are not mounted inadvertently. Longer cables must be allowed for in calculations. Their damping, and also that of damping elements, can be measured with a very high degree of accuracy using the signal generator and the power indicator at the desired frequency.

Suitable damping elements can be constructed as per Vieland (6) itself. Here I propose different dimensions, which are aimed at

maximising the structural return loss, $a_r$. At specific resistance values in the E24 range, shown in Fig.10, this happens automatically. The damping dimension is distorted, of course, which is no disadvantage here. Such damping elements can be connected up subsequent to the signal generator if you are not satisfied with the accuracy of the internal divider, or in order to weaken the noise source to a defined extent for the temperature method.

The difficulties increase at high frequencies. From values above 200 MHz, up to 500 MHz, you must very definitely apply the principles of assembly, as described by Vieland (6, 7).

It is very generally true that a structural element can no longer be thought of as concentrated if its dimensions exceed lambda/100. Pin-and-socket connections and cables will no longer be completely tight, which can be demonstrated by bringing them close to each other.
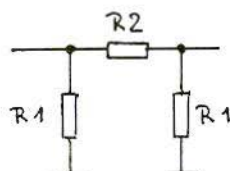
| R1/Ω  | 68    | 75    | 120  | 180  | 200  |
|-------|-------|-------|------|------|------|
| R2/Ω  | 160   | 120   | 51   | 30   | 27   |
| a/dB  | 16,33 | 13,98 | 7,71 | 4,96 | 4,44 |

$a_r = \infty$ für

$$\frac{R1^2 \cdot R2}{2R1 + R2} = Z_0^2 = (50\ \Omega)^2$$

**Fig.10: Damping factors for various values of resistors in the PI network shown above**

Sensitivity to manual contact (apart from heating influence) can be triggered by such leaks, but also by outside transmitters, conversion oscillators, digital gauges and standing waves (reflected power). In any case, it must be tuned out before any calibration whatever is undertaken.

Assistance can be obtained from improved screening, extremely short cables, and if necessary altering the measurement frequency or the conversion frequency.

# 5.
# SUMMARY

Three methods are proposed which make it possible to calibrate a noise source even in a "cellar laboratory".

The first is suitable for rapid at-a-glance calibration, the second is the standard method (if a sufficiently accurate signal source is available), and the third is based on physical principles and is independent of other standards. If the procedure is carefully carried out, high accuracy is obtained.

# 6.
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